Modeling, Characterization and Design of Line-of-Sight Wireless MIMO Channels

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Abstract

This doctoral thesis contains a collection of five papers preceded by an introduction. The papers investigate channel models for, design of, and performance analysis of wireless *multiple-input multiple-output* (MIMO) systems which are subject to a strong *line-of-sight* (LOS) channel component.

MIMO technology is embraced as one of the key technologies for fulfilling the demand for increased throughput and improved *quality of service* (QoS) in future wireless applications. This technology can both be employed to increase reliability, through diversity schemes such as e.g. maximum ratio combining and Alamouti coding, or to increase the spectral efficiency by spatial multiplexing schemes such as e.g. eigenmode transmission and V-BLAST. The performance of a wireless MIMO system is heavily dependent on the condition of the channel matrix, in the sense that the channel matrix should be of high rank for the MIMO system to achieve good performance. When the channel is such that the major part of the received power at the *receiver* (Rx) is due to multipath, fulfilling the high rank criteria is dependent on low correlation between the different subchannels. On the other hand, if the dominant component at the Rx is the deterministic LOS component, fulfilling the high rank criteria becomes dependent on the design of the two antenna arrays employed.

In this thesis we derive optimal antenna array designs for pure LOS channels with respect to *mutual information* (MI), when any combination of *uniform linear arrays* (ULAs) and *uniform planar arrays* (UPAs) are employed at the *transmitter* (Tx) and Rx. The important parameters with respect to design will be shown to be the antenna separation, antenna orientation, wavelength, transmission distance, and MIMO dimension. Moreover, we characterize the effects of these parameters deviating from their optimal values. The pure LOS channel matrix utilized is subsequently employed in a Ricean channel model also incorporating multipath, and performance is evaluated both analytically and numerically for different designs and multipath conditions. Furthermore, we investigate the performance of a possible future high frequency *fixed wireless access* (FWA) system based on the optimal design principle, to see how it works in a more realistic scenario. In general, the results show that a considerable gain is achieved if a design close to the optimal is possible for a MIMO system transmitting over a strong LOS channel.

The thesis also contains an analysis of the difference between the *spherical wave model* (SWM) and the *plane wave model* (PWM). The investigation is performed for systems utilizing ULAs, and it results in a framework that can be employed when evaluating when to apply the true SWM, and when the more simple approximate PWM gives sufficient modeling accuracy. Based on the framework, we conclude that the SWM should for example be applied in some typical WLAN scenarios.

Preface

This dissertation is submitted in partial fulfillment of the requirements for the degree of *philosophiae doctor* (PhD) at the Department of Informatics at the University of Oslo (UiO). My main supervisor for my PhD work has been Associate Professor Pål Orten at the University Graduate Center at Kjeller (UniK)/UiO, while my co-supervisor has been Professor Geir E. Øien at the Department of Electronics and Telecommunications at the Norwegian University of Science and Technology (NTNU).

The studies were carried out over a period of approximately three and a half years, from July 2003 to February 2007. In addition to the research, the PhD work includes one semester of course work, and one year of duty work at UniK outside Oslo. The duty work consisted mainly of teaching assistant responsibilities in different courses in telecommunication. The first year I mainly worked at the Department of Electronics and Telecommunications at NTNU, while the last two and a half years my workplace has been a combination of Nera Satcom (now Thrane & Thrane) at Billingstad outside Oslo and UniK.

The research was funded by Nera Networks (`http://www.nera.no/`) through the project *Low-Cost High-Capacity Terrestrial Radio* with support from the Research Council of Norway (NFR), while the assistantship was financed by UniK.

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which have made my last three and a half years very instructive and exiting.

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Furthermore, I would like to thank all the fantastic colleagues I have had the pleasure to get to know during my PhD studies. Both from the Department of Electronics and Telecommunications at NTNU, the wireless communication people at UniK, and the Research Department at Nera. You have made the PhD period an interesting and enjoyable experience. I would especially like to thank fellow PhD students Duc Van Duong, Sébastien de la Kethulle de Ryhove, Vegard Hassel, and Hans Jørgen Bang for both our scientific and social gatherings. My thanks also go to Vegard Hassel for reviewing the introduction of this thesis.

Finally, I would like to express gratitude to my family and friends. You have supported me and kept me in shape both socially and physically during the PhD work. You have always been there when I needed some time away from “channel models” and “Shannon capacity”, and for that I am very thankful.

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Contents

Abbreviations

3GPP	The 3rd generation partnership project
AP	Access point
ASP	Antenna separation product
AWGN	Additive white Gaussian noise
BER	Bit-error-rate
BS	Base station
CDF	Cumulative distribution function
CDMA	Code division multiple access
CSI	Channel state information
dB	Decibel
DoA	Direction-of-arrival
DoD	Direction-of-departure
EGC	Equal gain combining
EP	Equal power (power allocation)
ETSI	European Telecommunications Standards Institute
FWA	Fixed wireless access
GHz	Gigahertz
GSCM	Geometry-based stochastic channel model
HSDPA	High-speed downlink packet access
Hz	Hertz
IEEE	Institute of Electrical and Electronics Engineers
ISI	Inter-symbol-interference
LDPC	Low density parity check
LOS	Line-of-sight
MHz	Megahertz

MI	Mutual information
MIMO	Multiple-input multiple-output
MMSE	Minimum-mean-square error
MRC	Maximum ratio combining
MUD	Multiuser diversity
NLOS	Non-line-of-sight
OFDM	Orthogonal frequency division multiplexing
PDF	Probability distribution function
PMP	Point-to-multipoint
PTP	Point-to-point
PSK	Phase shift keying
PWM	Plane wave model
QAM	Quadrature amplitude modulation
QoS	Quality of service
RHS	Right hand side
RMD	Reflection/multiple diffraction
RT	Ray-tracing
Rx	Receiver
SC	Selection combining
SDMA	Space division multiple access
SIC	Successive interference canceler
SISO	Single-input single-output
SNIR	Signal-to-noise and-interference ratio
SNR	Signal-to-noise ratio
STBC	Space-time block code
STC	Space-time coding
STTC	Space-time trellis code
SU	Subscriber unit
SVD	singular value decomposition
SWM	Spherical wave model
TDD	Time division duplex
TDMA	Time division multiple access
Tx	Transmitter
ULA	Uniform linear array
UPA	Uniform planar array

V-BLAST	Vertical Bell Labs Space-Time Architecture
WF	Waterfilling (power allocation)
WLAN	Wireless local area network
WMAN	Wireless metropolitan area network

Part I

Introduction

Introduction

1 Background

1.1 Telecommunication trends

The telecommunication industry has evolved into one of the world's largest industries during the last decades. The main reasons for this development are the vast deployment of Internet and new telephone services. A tendency is that people are becoming more and more dependent on being connected, and to be able to send and receive information at any time from anywhere. This seems to be the case both at work and in our social lives.

During these last decades, communication has moved from the traditional analog domain to the digital domain. This makes sense for computer-to-computer communication, since the information is inherently digital. In addition, content such as voice, pictures, and videos, which are originally continues (in time, space, or amplitude), are usually represented digitally before transmitted. One advantage of this digitalization is that it makes it possible to benefit from advances in compression technology, which dramatically reduce the bit rate required while maintaining the perceptual quality.

A large portion of this information is transmitted through wired channels, e.g. cables and fibers, on its journey from the *transmitter* (Tx) to the *receiver* (Rx). However, this kind of wired communication requires expensive and time consuming deployment of cables, and is not very flexible with regards to changes in the network topology. Consequently, a considerable part of today's communication is done over wireless channels. For example, in the backbone of communication networks, radio relay systems facilitate flexible and fast deployment, for the last mile,[1] *fixed wireless access* (FWA) is a cost efficient alternative, and *base stations* (BSs) are employed in mobile phone systems to give the users mobility and flexibility.

1.2 The wireless channel

Moving from wired communications to wireless communications introduces a number of new challenges. First of all, the dynamic wireless channel is usually much more complicated than the more static wired channel (see e.g. [1–4] for a detailed description). Physical phenomena such as reflection, diffraction, and scattering cause what is referred to as *multipath*, i.e., multiple versions of the transmitted signal reach the Rx with different delays and amplitudes, and can thus be added either constructively or destructively. Furthermore, the wireless channel is subject to *shadowing*, which occurs when an obstacle is positioned between the Tx and Rx, resulting in a drop in the signal strength at the Rx. Both these phenomena lead to fluctuations in the signal strength received, which is referred to as *fading*. How rapid these fluctuations are depend on the speed of the Tx and Rx, and the speed of the objects causing the multipath and shadowing in the transmission environment.

In addition to these fluctuations, the mean power of the received signal will typically decrease with increasing transmission distance. This effect is referred to as *path loss*, and the rate of change is dependent on the transmission environment, e.g. urban, suburban, or open area. The received power is typically predicted to decay with the transmission distance raised to some power. For example, for free space propagation the received power decays with the transmission distance raised to the power of minus two [1, p.71].

Another important challenge for systems utilizing the wireless channel is *frequency reuse*. In contrast to wired communication, where for example frequency reuse in two neighboring cables is generally not a problem, the frequency usage in wireless communications must be planned carefully. To be able to reuse the frequency for wireless channels one must ensure that the interference from the co-channel users is at an acceptable level. Usually, this is achieved by assuring that there is sufficient geographical separation between the two systems using the same frequency, such that the path loss reduces the power of the interferers suffi-

[1] The last mile is often used in the communication industry to describe the final leg of delivering connectivity from a communications provider to a customer.

ciently. For this purpose the cellular concept is often introduced, where the total coverage area is split into cells utilizing different frequencies, and cells with sufficient separation are allowed to reuse the same frequency [1, Ch.2]. Moreover, frequency reuse can be further improved by reducing the beamwidth of the antennas employed, implying that the Tx transmits and the Rx receives interference from a smaller angle. In connection with cellular systems this is referred to as *sectoring*. To sum up, the wireless frequency spectrum is a very limited resource, and it is the regulatory authorities in each country that decide what systems that are allowed, and at which frequencies and power levels.

1.3 Spectral efficiency

Since the wireless frequency spectrum is such a scarce resource there is a strong desire to exploit it as efficiently as possible. Consequently, to follow the increasing demand on throughput in upcoming wireless applications, it is crucial to increase the number of bits per second transmitted per Hertz, i.e. increase the *spectral efficiency*. A lot of research effort has thus been put into different technologies which facilitate efficient frequency utilization over the years. In the following subsections we will give a brief description of three such areas of research.

1.3.1 Coding and modulation

An example of one such traditional topic is coding and modulation. The objective in this case is to find combinations of coding and symbol mapping that give reliable communication with as high spectral efficiency as possible subject to for example a given power constraint [5]. A big breakthrough in this discipline was made by Berrou et al. [6], with the sensational discovery of the turbo decoding principle in 1993. The word "turbo" is employed because the decoder utilizes a feedback loop just like a turbo-engine.

1.3.2 Dynamic transmission schemes

In recent years the attention of a lot of researchers has been turned towards the gain achieved by employing dynamic transmission schemes. For such schemes the transmission can adapt to changing channel conditions.

One such technology is *link adaptation* [7–10], where the channel quality is monitored through estimation and prediction, and made known to both the Tx and Rx. A transmission mode with high spectral efficiency is employed when the channel conditions are good, while a more robust transmission mode with a lower spectral efficiency is employed when the channel quality becomes worse. On the other hand, a traditional static transmission scheme must be designed to function for a worst case scenario, and does not have this possibility of changing spectral efficiency when the channel allows it. Consequently, the average spectral efficiency will typically be higher for dynamic systems compared to the traditional static systems. This increase in spectral efficiency is achieved at the expense of increased complexity and variable delay, since channel prediction/estimation, feedback, buffering, and multiple transmission schemes must be implemented.

Opportunistic scheduling is an example of another dynamic transmission scheme [11–13], which in contrast to the link adaptation scheme is only of interest for multiuser communications. The task of the scheduler is to distribute the system resources, e.g. timeslots in a *time division multiple access* (TDMA) scenario, between the different users in a smart way based on their channel condition. Users that experience favorable channel conditions, as measured by some suitable metric, are given priority to transmit or receive data. By giving priority to the users that have the best channel conditions, the system spectral efficiency can be increased. The gain obtained is referred to as *multiuser diversity* (MUD) gain. For such transmission schemes, the channel should preferably be rather dynamic, so all users get access to the channel within a certain timeframe, and properties such as *fairness* and *delay* are important design parameters.

1.3.3 Multiple antenna systems

Antenna arrays have been employed in wireless communication for some time. By distributing the antennas at the Tx or Rx in space, and thus utilizing the spatial domain, we can e.g. perform beamforming, interference rejection or diversity combining. By employing these techniques, which are referred to as *smart antenna* techniques, we can achieve increased *signal-to-noise ratio* (SNR), reduced interference, and a more reliable communication link [14, 15]. All these properties are favorable with respect to increasing spectral efficiency.

Multiple-input multiple-output (MIMO) can be viewed as an extension of the smart antenna techniques, and are based on combining multiple antennas both at the Tx and Rx. In addition to supporting traditional smart antenna techniques as mentioned above, MIMO systems have the potential of dramatically increasing the spectral efficiency by doing what is referred to as *spatial multiplexing* [16, 17]. Such systems have been extensively studied over the last years [18–20], and both commercial smart antenna and MIMO products are already hitting the market [21, 22].

2 MIMO systems

Results hinting at the potential of wireless MIMO systems was first published by Winters in [23]. The efficiency of wireless MIMO systems was further examined in some pioneering work by Telatar [16] and Foschini and Gans [17]. In these papers they showed that by employing multiple antennas at the Tx and Rx, and combining them in a smart way, the spectral efficiency will increase dramatically. Actually, for rich scattering transmission environments the spectral efficiency will theoretically increases linearly with the minimum number of antennas at the Tx and Rx [24, p.341]. Another interesting property of MIMO communication is that the multipath from the transmission environment, which usually is a problem for the communication link, is utilized and can be turned into a benefit [25]. These findings sparked the interest of many researchers, and a tremendous amount of research papers have been published over the last decade on such diverse topics as information theory, channel modeling, Tx and Rx structures, channel estimation/prediction, feedback etc. The increase in performance promised by MIMO technology is achieved at the expense of complexity and hardware cost.

An example of a single link narrowband MIMO system is illustrated in Figure 1.1. Here, the number of Tx antennas is denoted N while the number of Rx antennas is denoted M. The channel, which in the case of MIMO transmission is described by a matrix, is given by $\sqrt{\chi} \cdot \mathbf{H}$. Here, χ is the common power gain over the channel, while $\mathbf{H}$ is the normalized channel matrix linking the N Tx antennas with the M Rx antennas. The element in row i and column j of the channel matrix is denoted by $h_{i,j}$. It is further assumed that $h_{i,j} \in \mathbb{C}$. This assumption will be justified in the next section, while the normalization implies $E[h_{i,j}h_{i,j}^*] = 1$, where $h_{i,j}^*$ denotes the complex conjugate of $h_{i,j}$ and $E[\cdot]$ is the expectation oper-

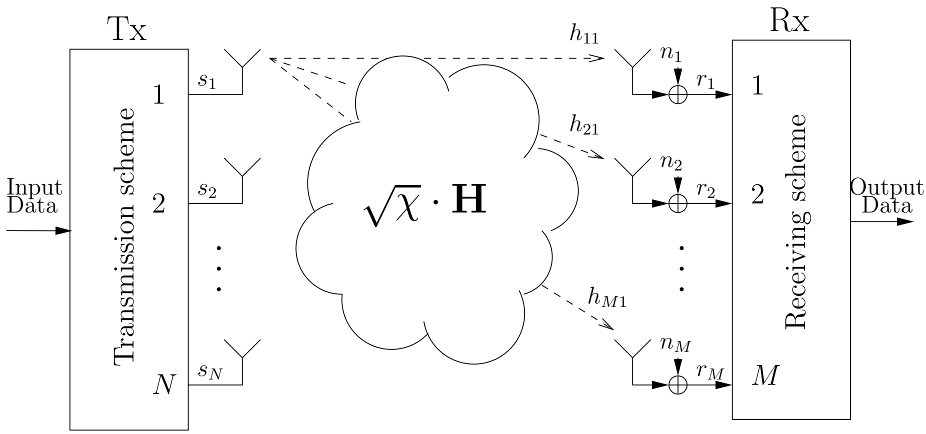

FIGURE 1.1: An illustration of a general single link narrowband MIMO system with N Tx antennas and M Rx antennas.

ator. When assuming an *additive white Gaussian noise* (AWGN) channel, the narrowband MIMO transmission can be expressed in the complex baseband domain as [18, p.53]

$$\mathbf{r} = \sqrt{\chi} \cdot \mathbf{Hs} + \mathbf{n}, \tag{1.1}$$

where $\mathbf{r} \in \mathbb{C}^{M\times 1}$ is the received signal vector, $\mathbf{s} \in \mathbb{C}^{N\times 1}$ is the transmitted signal vector, and $\mathbf{n} \in \mathbb{C}^{M\times 1}$ is the AWGN vector, i.e. [2] $\mathbf{n} \sim \mathcal{CN}(\mathbf{0}_{M\times 1}, \sigma_n^2 \cdot \mathbf{I}_M)$, where σ_n^2 is the noise power at each Rx antenna and $\mathbf{I}_M$ an identity matrix of dimension $M \times M$. Consequently, the signal received at each Rx antenna is a sum of different versions (dependent on the subchannel gain) of all the transmitted signals plus noise.

We refer to the transmission scheme used by the MIMO system as *space-time coding* (STC), as it has the possibility to code over both the space and time dimension. STC schemes can both be designed to maximize throughput by employing *spatial multiplexing*, and to increase reliability by employing diversity schemes. Some STC schemes also try to combine these two properties, and they result in some kind of compromise between increased throughput and reliability [26].

[2] $\mathcal{CN}(\mathbf{x}, \mathbf{Y})$ denotes a circularly symmetric complex Gaussian distributed random vector, with mean vector $\mathbf{x}$ and covariance matrix $\mathbf{Y}$.

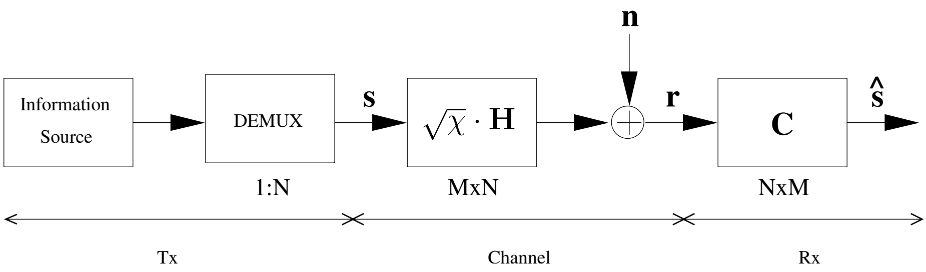

FIGURE 1.2: A MIMO system employing spatial multiplexing with a linear Rx scheme.

2.1 Spatial multiplexing

As an example of how MIMO transmission works, we will explain the principle of spatial multiplexing in more detail. A spatial multiplexing system is illustrated in Figure 1.2. Here, the incoming symbols from the information source are demultiplexed to the different antennas, i.e. different symbols are transmitted at each antenna. Consequently, the symbol rate is $1/N$ on each Tx branch compared to what is delivered from the information source. This transmission architecture is often referred to as *Vertical Bell Labs Space-Time Architecture* (V-BLAST) in the literature [27]. In the figure we have also assumed that a linear detector is employed, indicating that we multiply $\mathbf{r}$ with a matrix $\mathbf{C}$ to find our estimate of the transmitted signal $\hat{\mathbf{s}}$. For simplicity it is assumed that $N \leq M$, i.e. we do not transmit more symbols simultaneously than there are Rx antennas. At the Rx we now want to find an estimation of $\mathbf{s}$ based on $\mathbf{r}$.

If we first assume that the noise is negligible, we observe that (1.1) is reduced to M equations with N unknowns. From linear algebra we know that it is possible to find the unknowns if the rank of $\mathbf{H}$ is at least as large as the number of unknowns. We find our estimate $\hat{\mathbf{s}}$ by multiplying $\mathbf{r}$ with a matrix $\mathbf{C}$ given by the *Moore-Penrose pseudoinverse* [28, p.490], i.e.

$$\mathbf{C} = \frac{1}{\sqrt{\chi}}(\mathbf{H}^H\mathbf{H})^{-1}\mathbf{H}^H, \tag{1.2}$$

where $\mathbf{H}^H$ denotes the Hermitian transpose of $\mathbf{H}$, and the detection scheme is referred to as *zero-forcing* since it removes all the interference between the different symbols transmitted. Based on this example we can make some general comments: The rank of $\mathbf{H}$ is a very important property for MIMO systems. It is equal to the degrees of freedom offered by the MIMO channel and thus the number of possible spatial channels available. Furthermore, the MIMO transmission requires channel state information at least at the Rx. This channel state information can be obtained in different ways. One procedure is based on known symbols (pilot symbols), which is used to predict/estimate the subchannels at the Rx [29, 30]. Quite a bit of research have also been done on blind techniques, where no explicit training signals are used, instead the Rx estimates the channels based on the data transmitted [31, 32].

If we now increase the noise power in (1.1), the zero-forcing algorithm is not necessary a smart way of detecting the signal anymore, since it can lead to noise enhancement. Another Rx candidate, which is also linear, could then be the *minimum-mean-square error* (MMSE) detector which performs an optimal trade-off between the AWGN and interference from the other signals. If we assume that the total Tx power P is divided equally between the elements of $\mathbf{s}$, the MMSE detection is given by [28, p.495]

$$\mathbf{C} = \frac{1}{\sqrt{\chi}} \left(\mathbf{H}^H \mathbf{H} + \frac{N \cdot \sigma_n^2}{\chi \cdot P} \cdot \mathbf{I}_N \right)^{-1} \mathbf{H}^H. \tag{1.3}$$

We observe that when the SNR increases, the second term in the parentheses decreases, and $\mathbf{C}$ from (1.3) approaches (1.2) as expected. On the other hand, if the SNR becomes small, the second term in the parentheses becomes dominant, consequently the receiver approaches a matched filter receiver [28, p.489].

There also exist *non-linear* detection algorithms, which in general perform better than the linear detectors. An example of such a detector is the *successive interference canceler* (SIC) [28, p.499]. The SIC is based on detecting the symbols one at a time, and then subtracting them from the received vector to reduce interference when detecting the remaining signals. For this algorithm, properties such as *ordering* and *error propagation* are important with respect to performance.

Actually, the detection algorithms described above are not invented mainly for MIMO systems, but are well known detection algorithms for signals subject to interference and noise. For example, one will find versions of the zero forcer, MMSE algorithm, and SIC in equalizer design, where the interference is between consecutive symbols in time (*inter-symbol-interference* (ISI)) [33, Ch.10], and in *code division multiple access* (CDMA) receivers, where the interference is between the different user signals [33, Ch.15].

If channel state information is fed back to the Tx, the transmission scheme can be made more sophisticated. One possibility is then to transmit on the eigenmodes of the channels, i.e. transmitting along the eigenvectors of the channel matrix. This creates a number of parallel orthogonal subchannels equal to the rank of $\mathbf{H}$ [34].

2.2 Diversity schemes

As mentioned above, utilizing the spatial dimension by employing more than one antenna at the Tx and/or Rx also makes coding for increased link quality feasible. This is done by increasing the *diversity order*[3], which makes the *bit-error-rate* (BER) versus SNR curve steeper. Different Rx diversity schemes, such as e.g. *maximum ratio combining* (MRC), *selection combining* (SC), *equal gain combining* (EGC), and *switched combining*, have been studied for many years [2, Ch.6].

Transmission schemes resulting in Tx diversity on the other hand, is a more recent topic of research. The consequence of introducing redundancy (in time and space) between the signals transmitted, is that the throughput goes down while the diversity order is increased. Such STC schemes can be divided into two main categories; *space-time trellis codes* (STTCs) and *space-time block codes* (STBCs) [19]. The key development of STC was revealed in [35], where a STTC scheme was suggested which required a multidimensional Viterbi detector at the Rx. These codes provide a Tx diversity order equal to the number of transmit antennas, but require a relatively complex Rx algorithm. The interest in STC increased further with the discovery of STBCs. These codes have the advantage of allowing simple linear Rx structures due to the design of the codes. The best known STBC is probably the so-called Alamouti code, named after its inventor [36]. This scheme utilizes two antennas, and by coding two information symbols over two time intervals, it achieves full transmit diversity. The maximum diversity order attainable for the whole multiple antenna system is $M \cdot N$.

[3] The impact of order-m diversity is to raise the BER without diversity to the power of m [28, p.543].

3 The wireless MIMO channel: Characterization and modeling

The large performance gain initially promised by wireless MIMO technology was derived based on the assumption of *independent and identically distributed* (i.i.d.) complex Gaussian subchannels [16, 17]. In this case, all elements in the channel matrix are independently distributed and therefore uncorrelated. For this to be true in a real communication scenario, we need a rich scattering transmission environment where a sufficient number of multipath components are received at the Rx, while the antennas should be separated sufficiently to avoid spatial correlation [37, 38]. This is of course not necessarily true for all transmission scenarios. Since the performance is heavily dependent on the MIMO channel, and to get a better insight into the actual performance achieved, it is important to find more accurate models valid for these other scenarios as well. At the same time it is desirable that the model is easy to use and therefore not too complex. Consequently, when choosing a channel model we are often constrained by the trade-off between model accuracy and model simplicity. We will start by describing some properties of the wireless channel in general, before we turn our attention to modeling of the wireless MIMO channel.

3.1 Time and frequency characterization

As mentioned earlier, transmitting over a wireless channel often results in multipath propagation. When the delays between the different received components are large, and the power is strong enough to considerably influence the received amplitude and/or phase of the received signal, we characterize the fading as *frequency-selective*. By taking the *Fourier transform* of the impulse response of such a channel, we observe that the frequency response is not flat over the whole signal bandwidth, this being the rationale behind the name frequency-selective fading. The parameter employed to characterize the "flatness" of the channel is the *coherence bandwidth*, denoted by B_c. Two sinusoids that are separated in frequency with less than B_c are by definition strongly correlated. The coherence bandwidth can be calculated based on the *power delay profile*, and the procedure is described in [1, Ch.4] and [2, Ch.2]. Channels that are not frequency-flat in the first place, can be split into several frequency-flat subchannels by employing multicarrier transmission such as e.g. *or-*

thogonal frequency division multiplexing (OFDM) [39, 40]. Alternatively the frequency-selectivity can be compensated for by employing an equalizer at the Rx [33, Ch.10]. The distinction between frequency-selective and frequency-flat fading is also often referred to as broadband and narrowband communications respectively.

Usually, the properties of the wireless channel will vary with time. This change is generally caused by either relative movement between the Tx and Rx, or by movement of objects in the transmission environment. To characterize the speed of this channel variation we employ the *coherence time*, denoted by T_c. Coherence time is a statistical measure of the time duration over which the channel impulse response is essentially invariant, and thus quantifies the channel response similarity at different times. Consequently, if the symbol period is smaller than T_c, the channel response is approximately constant during the transmission of the symbol, and the fading in this case is referred to as *slowly varying*. On the other hand, if the symbol period is larger than T_c, one should assume that the channel response may change during the transmission of the symbol, and this kind of fading is referred to as *fast fading*. The coherence time can be calculated based on the *Doppler shift*, and the procedure is described in [1, Ch.4] and [2, Ch.2].

In Figure 1.3 we have illustrated the principle of characterizing the channel fading as described above. The two dimensions of the matrix in the figure are the symbol period, denoted by T_s, and the signal bandwidth, denoted by B_s.

In the system model given in (1.1) it was assumed that the elements of $\mathbf{H}$, i.e. $h_{i,j}$, were given by complex constants. This is true when the channel is slowly varying ($h_{i,j}$ is constant during transmission of one $\mathbf{s}$) and frequency-flat ($h_{i,j}$ is the same for the whole signal bandwidth). Consequently, the system model in (1.1), and thus the system investigated in this thesis, belongs in the shaded region in the lower left corner of Figure 1.3. One common way of modeling a slowly varying fading channel is by assuming that $\mathbf{H}$ is constant over $K \in \mathbb{Z}$ consecutive symbol periods at a time, and then after K symbol periods a new realization of $\mathbf{H}$ is valid. This is often referred to as a *block fading channel* in the literature [41, 42].

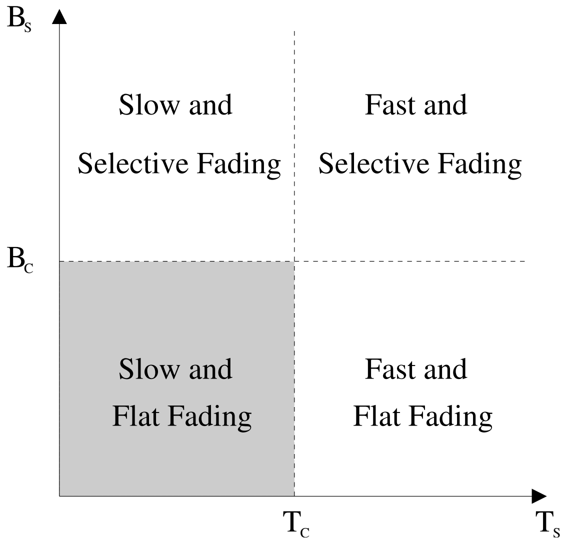

FIGURE 1.3: A matrix characterizing combinations of fading as a function of symbol period (T_s) and signal bandwidth (B_s).

3.2 Modeling of H

When modeling the channel matrix, we follow the same procedure as in [4], and distinguish between *physical channel models* and *analytical channel models*.

3.2.1 Physical channel models

Physical models focus on the double-directional propagation mechanisms between the location of the Tx and Rx without taking into account the specific antenna configuration. Double-directional in this context means that the model includes angular information at both ends of the communication link, i.e. *direction-of-departure* (DoD) and *direction-of-arrival* (DoA) [43]. To find the channel matrix **H** (transfer function), we need to combine the physical model with a realization of the Tx and Rx antenna array configuration. This way of modeling the channel has the advantage that different antenna array configurations can be evaluated for the same transmission environment, without the need to analyze the transmission environment for each configuration.

We subdivide the physical channel models into two main categories; *deterministic models* and *stochastic models*. As the name states, the stochastic physical channel models combine stochastic models (typically derived based on measurements [44]) with physical laws of propagation when searching for the channel characteristics. There are two types of such models, *geometry-based stochastic channel models* (GSCM) [45–47] and *non-geometric stochastic channel models* [48–50].

Physical propagation models are referred to as deterministic if they try to reproduce the actual physical radio propagation for a given specific environment. The advantage of this kind of modeling is that it allows for very high accuracy. Actually, due to this high accuracy, this method may be employed instead of measurements in some cases, because setting up and carrying out a measurement campaign can be both time consuming and difficult. To model the transmission, the geometrical and electromagnetic characteristics of the transmission environment can be stored in so-called environment databases, while techniques such as for example *ray-tracing* (RT) [51, Ch.12] can be used to simulate the electromagnetic propagation process. In the work presented in this thesis we will utilize a simple version of the deterministic physical channel model, equivalent to the one applied in [52].

3.2.2 Analytical channel models

In contrast to physical channel models, analytical channel models search to characterize the channel matrix $\mathbf{H}$ in a mathematical way without explicitly accounting for the electromagnetic wave propagation [4]. These models are popular in performance analysis and in MIMO algorithm design, because of their often simpler construction. The analytical channel models can be further split into two classes [4]; *correlation-based analytical models* and *propagation-motivated analytical models*. The latter group of models characterizes the MIMO matrix based on propagation parameters [53, 54], while the first group of models characterizes the MIMO matrix statistically with focus on the correlation between the subchannels. Since the correlation model will be employed in this thesis, it will be described in more detail in the reminder of this section.

When employing correlation-based analytical models for frequency-flat fading channels, it is common to assume that the channel gains of the subchannels follow a multivariate complex Gaussian distribution. The channel matrix for the system model in (1.1) then becomes [18, p.41]

$$\mathbf{H} = \sqrt{\frac{K}{1+K}} \cdot \mathbf{H}_{\mathrm{LOS}} + \sqrt{\frac{1}{1+K}} \cdot \mathbf{H}_{\mathrm{NLOS}}, \tag{1.4}$$

where $\mathbf{H}_{\text{LOS}} \in \mathbb{C}^{M\times N}$ is the deterministic *line-of-sight* (LOS) channel matrix, $\mathbf{H}_{\text{NLOS}} \in \mathbb{C}^{M\times N}$ is the stochastic zero mean *non-LOS* (NLOS) channel matrix, and K is the Ricean K-factor defined as the ratio between the powers of the LOS and NLOS channel components. Due to the normalization of $\mathbf{H}$ introduced in (1.1), each element in $\mathbf{H}_{\text{LOS}}$ and $\mathbf{H}_{\text{NLOS}}$ should have a power equal to one. This channel is often referred to as a Ricean channel, or alternatively when $K = 0$, a Rayleigh channel. The channel matrix in (1.4) has the property $E[\mathbf{H}] = \mathbf{H}_{\text{LOS}}$. The elements of $\mathbf{H}_{\text{LOS}}$ may often be determined by simple analytical models, usually assuming that $\text{rank}(\mathbf{H}_{\text{LOS}}) \approx 1$, as e.g. in [55, 56]. However, for some scenarios one should employ more correct models, and thus take the transmission environment into account, to satisfy the required modeling accuracy [52, 57, 58].

We now consider the stochastic NLOS part of the channel matrix, and define $\mathbf{h}_{\text{NLOS}} \stackrel{\Delta}{=} \text{vec}(\mathbf{H}_{\text{NLOS}})$, where $\text{vec}(\cdot)$ is the vectorization operator, which stack the columns of the matrix in the argument on top of each other. The distribution of $\mathbf{h}_{\text{NLOS}}$ is then given by [59, p.43]

$$f(\mathbf{h}_{\text{NLOS}}) = \frac{1}{\pi^{MN}\det(\mathbf{R}_{\text{NLOS}})}\exp\left(-\mathbf{h}_{\text{NLOS}}^{H}\mathbf{R}_{\text{NLOS}}^{-1}\mathbf{h}_{\text{NLOS}}\right), \qquad (1.5)$$

where $\det(\cdot)$ is the matrix determinant operator, and

$$\mathbf{R}_{\text{NLOS}} = E[\mathbf{h}_{\text{NLOS}}\mathbf{h}_{\text{NLOS}}^{H}] \qquad (1.6)$$

is the correlation matrix. Consequently, $\mathbf{R}_{\text{NLOS}} \in \mathbb{C}^{MN\times MN}$ contains information on the correlation between all the subchannels. The normalization mentioned above gives the constraint $(\mathbf{R}_{\text{NLOS}})_{i,i} = 1$, where $(\cdot)_{i,j}$ denotes the element in row i and column j. Different approaches to modeling this correlation matrix have been suggested in the literature. The simplest model assumes i.i.d. subchannels, and in this case we get $\mathbf{R}_{\text{NLOS}} = \mathbf{I}_{MN}$. For this to be true in a real transmission environment, the channel must be rich scattering, and the antenna elements must be sufficiently spatially separated at both Tx and Rx [37, 38]. Another popular model is the Kronecker model [18, p.40], where it is assumed that the correlation at the Tx and Rx can be separated. The correlation matrix in this case can be written as a Kronecker product, i.e.

$$\mathbf{R}_{\text{NLOS}} = \mathbf{R}_{\text{Tx}} \otimes \mathbf{R}_{\text{Rx}}, \qquad (1.7)$$

where the elements of $\mathbf{R}_{\mathrm{Tx}}$ and $\mathbf{R}_{\mathrm{Rx}}$ are the correlation between the antennas at Tx and Rx respectively. The Kronecker model has become popular because of its simplicity, however when considerable correlation exist, measurements show that the model may underestimate the actual channel capacity [60]. A third correlation model found in the literature is the so-called Weichselberger model [61]. By introducing a coupling matrix in the correlation model, the Weichselberger model aims at removing the restriction of separable Tx and Rx correlation given by the Kronecker model.

4 MIMO channel capacity

One way of investigating the potential performance of a communication system is to look at the *channel capacity*. The channel capacity is by definition the maximum information rate that a channel can support with arbitrary small error probability [62, p.194]. The quantity is found by maximizing the *mutual information* (MI) between the input and output of the channel over all possible input distributions which adhere to the transmission constraints (e.g. power constraints). The channel capacity for AWGN channels was first derived by Claude Shannon in his historical paper from 1948, *A mathematical theory of communications* [63].

Even though the channel capacity quantifies the potential for information transmission over the channel, and gives us some hints as to which properties are important when chasing this upper bound (e.g. long block lengths and random codes), it does not tell us explicitly how to design a transmission scheme that achieves this rate. Consequently, the channel capacity is often employed as a benchmark, to evaluate how far a transmission scheme is from the theoretically optimal performance.

We now turn our attention to MIMO systems, more precisely the transmission over the slow and flat fading AWGN channel as described in (1.1). This is the same system as studied in [16, 17], and by assuming $\mathbf{s} \sim \mathcal{CN}(\mathbf{0}_{N\times 1}, \mathbf{Q})$, the MI between $\mathbf{s}$ and $\mathbf{r}$ (for a given $\mathbf{H}$) is given by

$$\mathcal{I} = \log_2 \det\left(\mathbf{I}_M + \frac{\chi}{\sigma_n^2} \cdot \mathbf{HQH}^H\right) \quad \text{bits/s/Hz}. \tag{1.8}$$

Here it is assumed that we transmit 1 complex symbol per Hertz, which corresponds to the maximum allowed rate without ISI for a passband channel (Nyquist criteria [33, p.556]). The base-2 logarithm is employed to get the result in bits (information bits per channel use). Actually, if the AWGN MIMO transmission described in (1.1) is subject to an average power constraint, the circularly symmetric complex Gaussian $\mathbf{s}$ applied in (1.8) is what maximizes the MI [16].

If we now further presume that the total available Tx power for each transmitted $\mathbf{s}$ is P, there exist several different strategies on how to distribute this power between the different transmitted symbols. This choice affects (1.8) through the covariance matrix $\mathbf{Q} = E[\mathbf{s}\mathbf{s}^H]$. Two different strategies will now be described, i.e. *waterfilling* (WF) power allocation and *equal power* (EP) allocation.

4.1 Waterfilling (WF) power allocation

WF power allocation for MIMO systems is implemented by first creating individual subchannels, and then by allocating power according to the WF principle based on the channel gains on these different subchannels. The $\mathbf{Q}$ in this case becomes [16]

$$\mathbf{Q} = \mathbf{V}\tilde{\mathbf{Q}}\mathbf{V}^H, \tag{1.9}$$

where $\mathbf{V}$ is a unitary matrix where the columns are equal to the eigenvectors of $\mathbf{H}$, i.e. transmission is performed along the eigenvectors of the channel matrix. This transmission technique is often referred to as eigenmode or eigenbeam transmission. $\tilde{\mathbf{Q}}$ is a diagonal power allocation matrix, where $(\tilde{\mathbf{Q}})_{i,i}$ is equal to the power allocated to subchannel i, subject to the constraint $\sum_i (\tilde{\mathbf{Q}})_{i,i} \leq P$. The elements of this matrix are found by the following relation [16]:

$$(\tilde{\mathbf{Q}})_{i,i} = \left(\xi - \frac{\sigma_n^2}{\chi \cdot \sigma_i^2}\right)_+, \tag{1.10}$$

where σ_i is the ith singular value of $\mathbf{H}$, ξ is the water level found from the power constraint $\sum_i (\tilde{\mathbf{Q}})_{i,i} = P$, and $x_+ \stackrel{\Delta}{=} \max(0, x)$.

Actually, it can be shown that this transmission scheme is what maximizes the MI in the general case [16], and thus achieves the channel capacity. It is important to note that this transmission scheme requires knowledge of the channel at the Tx, because both $\mathbf{V}$ and σ_i are dependent on $\mathbf{H}$. For a symmetric two-way *time division duplex* (TDD) system, this could be realized by estimating/predicting the reciprocal channel parameters on both sides of the link, otherwise a feedback loop from the Rx (where the channel is estimated/predicted) to the Tx is required.

4.2 Equal power (EP) allocation

Another power allocation strategy frequently employed is the EP allocation. As the name states, it is based on equally dividing the available power between the N transmit antennas. The $\mathbf{Q}$ in this case becomes [16, 17]

$$\mathbf{Q} = \frac{P}{N} \cdot \mathbf{I}_N. \tag{1.11}$$

The EP allocation scheme is reasonable for systems that lack channel state information at the Tx. Moreover, this transmission scheme is shown to be optimal for an uncorrelated Rayleigh channel [16], i.e. the channel in (1.4) with $K = 0$ and $\mathbf{R}_{\text{NLOS}} = \mathbf{I}_{MN}$, when the channel is not known at the Tx. Consequently, the obtained MI is the channel capacity in this case. On the other hand, when $K \neq 0$, the MI obtained by EP transmission is not necessarily the channel capacity any more [64]. In [65] it was shown that one should transmit along the eigenvectors of the mean matrix (i.e. $\mathbf{H}_{\text{LOS}}$) in this case. Consequently, we use the same strategy as e.g. in [66], and even though we do not assume channel state information at Tx, the resulting MI is not referred to as capacity, because we know that there still exists a better covariance matrix than the one given in (1.11) resulting from EP allocation.

4.3 MI for fading channels

When the channel matrix is modeled as stochastic, as suggested by some of the models in Section 3, the MI given in (1.8) becomes a random variable. There are different ways of characterizing the MI for such a channel.

When the channel is ergodic, i.e. transmission time is long enough to reveal the ergodic properties of the channel, the average MI is a meaningful channel characteristic [16, 67]. The average MI is found by averaging the MI from (1.8) over all possible channel matrix realizations, i.e.[4] $\bar{\mathcal{I}} = E[\mathcal{I}]$.

On the other hand, when the channel does not significantly change during the duration of the total transmission time, there may be a non-zero probability that the value of the actual transmitted rate, no matter how small, exceeds the instantaneous MI [16, 67]. In this situation, we

[4] This is referred to as the *ergodic capacity* for transmission schemes that are capacity achieving.

should also take into account the distribution of the MI which in turn depends on the channel. An interesting channel characteristic in this case is the probability that the instantaneous MI falls below a given threshold, which is given by the MI *cumulative distribution function* (CDF), i.e.[5] $F_{\mathcal{I}}(\nu) = \Pr\ [\ \mathcal{I}\ <\ \nu\ ]$.

A considerable amount of papers have been published on MI and capacity results for different MIMO channels. Most of the work on this topic uses a variant of the correlation-based analytical channel (Ricean channel) described in (1.4) [68]. The seminal papers by Telatar [16] and Foschini and Gans [17], treated the uncorrelated Rayleigh channel. One of the major findings in this work was that the capacity of such channels scales linearly with $\min(M, N)$. Examples of other channels that have been studied extensively are the correlated Rayleigh channel employing different correlation models [69–71] and the general Ricean channel [55, 56, 66, 71–74].

5 Fixed wireless systems

In wireless communications it is common to distinguish between fixed and mobile systems. Sometimes, it is reasonable to define a third type of communication as well, so called nomadic communication, which implies that the user is fixed during communication, but can change position when not active. An example of a typical nomadic network is a *wireless local area network* (WLAN). In this work we will not distinguish between fixed and nomadic networks, because we are only interested in whether there is movement during communication or not. Consequently, for fixed systems the user terminal is assumed to be at a fixed position and have a fixed orientation when communicating. Even though the users are assumed fixed, objects in the transmission environment may change position, thus causing time varying fading as discussed in Section 3. Mobile systems on the other hand should, as the name states, enable mobility for the users during operation. Mobile systems are often more complex to design than their fixed counterpart, because they are often subject to much faster fading. In addition, new advanced algorithms are required to facilitate for example *handoff* between neighboring cells for cellular networks [2, Ch.12]. One consequence of

[5] This is referred to as the *outage capacity* for transmission schemes that are capacity achieving.

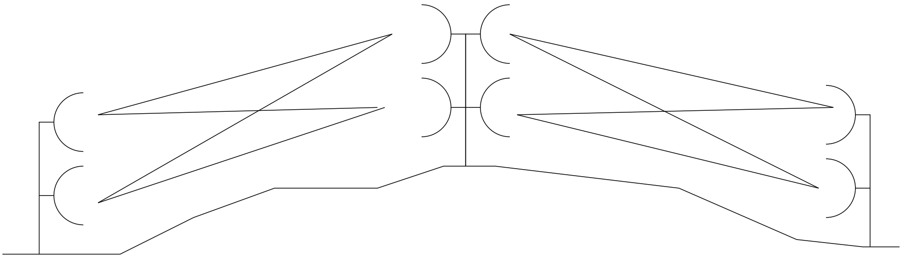

FIGURE 1.4: Illustration of a PTP network with two links to reach from the source to the destination. Both links are 2×2 MIMO systems.

this complexity difference is that fixed systems are often designed and able to give a larger throughput than mobile systems. However, we should also mentioned that fixed systems often are designed for higher frequencies, which introduce new challenging channel effects such as for example rain attenuation and scintillation.

In this section we will concentrate on fixed systems as defined above, because it is for these systems that the results given in the subsequent chapters are most relevant. Of course MIMO technology in general is very relevant for mobile systems, as can e.g. be seen in the standardization work done by *The 3rd generation partnership project* (3GPP) [75] on *high-speed packet access* (HSPA) [76] and by the *Institute of electrical and electronics engineers* (IEEE) on 802.16e [77]. However, the results presented in this thesis are mainly relevant for fixed communication, and in the following two subsections, we will briefly describe two subgroups of fixed networks; *point-to-point* (PTP) networks and *point-to-multipoint* (PMP) networks.

5.1 Point-to-point networks

PTP networks consist of one or more fixed PTP communication links. An example of such a system is illustrated if Figure 1.4, where two links are employed to reach from the source to the destination.

One type of such systems are *radio relay systems*, operating e.g. in the backbone of a communication network. These systems utilize the fact that they can be designed to have a strong LOS component between the Tx and Rx to increase their transmission rate. Radio relay systems also employ highly directive antennas, which limit the multipath and in-

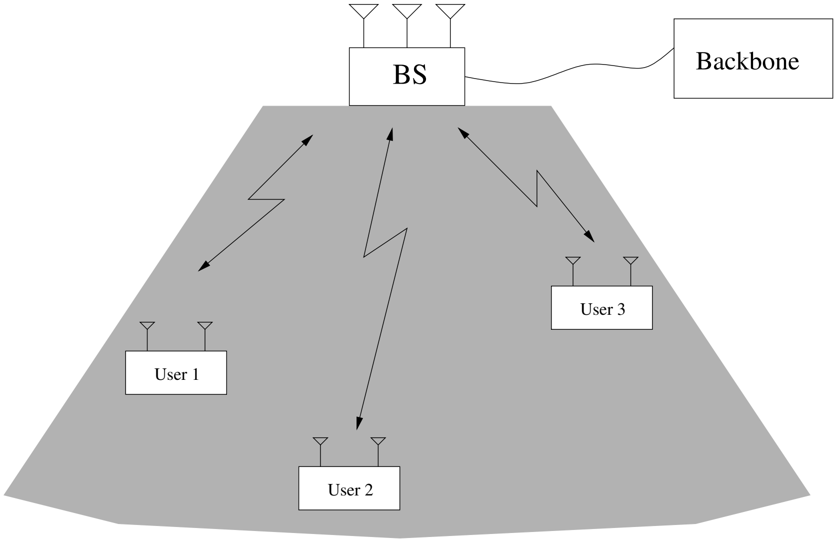

FIGURE 1.5: Illustration of a PMP network containing a BS with three antennas and three users with two antennas each. The shaded area illustrates the coverage area.

crease the link quality. A single link operates typically over a distance of 20-60 km. Even though MIMO technology for these systems has not yet been extensively studied, smart antenna techniques such as spatial diversity are widely deployed in today's radio relay systems. More details on design issues for such systems can be found in [78–82].

5.2 Point-to-multipoint networks

In a PMP network there is a centralized unit, often referred to as a BS or *access point* (AP), that consists of one or more broad beam antennas that should cover the whole area containing the relevant users, i.e. the coverage area. This centralized unit is connected to the rest of the network, either wired e.g. by a fiber cable, or wireless with e.g. a radio relay link. An example of a PMP network is illustrated in Figure 1.5. When a PMP network is to be employed for a large area, e.g. a city, it is common to employ several BSs to achieve sufficient coverage. Careful planning is then required to ensure that different BS systems do not interfere too much with each other.

We can distinguish between two different types of PMP networks depending on their mode of operation; *LOS networks* and *NLOS networks*. The LOS networks are networks where each communication link requires a LOS component to function properly [83]. In this case the user terminals are often installed by engineers, and directional antennas are employed and oriented by a technician with a LOS to the BS. Typical properties for such systems are high frequencies and/or long range. One example of such a network is the *wireless metropolitan area network* (WMAN), which is a typical last mile technology. There are several standards available, such as for example the IEEE 802.16 family [84], HIPERACCESS [85], and HIPERMAN [86] (note that some variants of these standards also support NLOS communication).

The user terminals of NLOS systems on the other hand, are typically arbitrarily positioned at the convenience of the end user, for example inside an office or house. Often the location of these user terminals are such that no obstruction-free LOS view of the BS antenna is available. These systems may be exposed to very large variations in attenuation and fading because they should cope with both LOS and NLOS conditions. This fact presents new challenges to system designers in their efforts to provide reliable high-speed communication. A typical example of such networks are WLANs, where the most well known standards are probably the 802.11 family [87] developed by the IEEE, and HIPERLAN/2 [88] developed by ETSI. Actually, MIMO technology is already included in ongoing standardization for these systems, i.e. IEEE 802.11n, where products based on a draft of this standard can be found in stores today [89–91].

6 Scope of the thesis

As mentioned in the preface, this PhD work has been part of an industry project at *Nera Networks*. Their product portfolio contains different high frequency (typically in the region 5–40 GHz) fixed wireless communication systems, i.e. both fixed wireless access systems and radio relay systems. Most research on wireless MIMO systems done in the last decade have focused on the MIMO gain given by uncorrelated subchannels resulting from multipath propagation. However, the motivation behind this PhD work was to explore the possibility of utilizing wireless MIMO technology for channels with a strong LOS component and limited multipath, as is the case for most systems developed by Nera Networks.

Before we present the contributions of each of the included papers in the next section, some general properties valid for all the papers will be summarized in the reminder of this section. The work presented in the thesis is based on the transmission model in (1.1), and thus the channel investigated belongs to the shaded region in Figure 1.3. Furthermore, in the consecutive chapters, the channel matrix is always modeled by a version of the correlation-based model as described in Section 3.2.2. The stochastic NLOS part of the channel matrix, i.e. $\mathbf{H}_{\text{NLOS}}$, is either modeled with no correlation at all, or with the Kronecker correlation model as described in (1.7). The deterministic LOS part of the channel matrix, i.e. $\mathbf{H}_{\text{LOS}}$, is found by using a simple form of the deterministic physical channel model as described in Section 3.2.1, where only the direct LOS component between the Tx and Rx is considered. The antenna arrays investigated in the thesis are either *uniform linear arrays* (ULAs) or *uniform planar arrays* (UPAs).

In all the included papers in this thesis the MI is employed as performance measure in one way or another. Both the case of WF power allocation along the eigenvectors of the channel matrix (Section 4.1), which is optimal when the channel is known at the Tx, and the case of EP allocation (Section 4.2), which is reasonable when the channel is only known at the Rx, are treated.

7 Contributions of the included papers

This thesis consists of five papers numbered with the capital letters A–E. In this section we will present a brief summary of these papers.

7.1 Paper A

Frode Bøhagen, Pål Orten, and Geir E. Øien, "Design of Capacity-Optimal High-Rank Line-of-Sight MIMO Channels," *Research Report 352*, Department of informatics, University of Oslo, ISBN 82-7368-309-5, ISSN 0806-3036, March 2007. *Available at: http://www.digbib.uio.no/*

In Paper A we give a more comprehensive presentation of the results we first presented in [57, 92]. The paper gives a procedure on how to achieve a high rank channel matrix, and thus high MIMO channel capacity, for pure LOS channels. The main difference between Paper A and [57] is that Paper A utilizes an alternative procedure to derive the results, which gives more insight into the solution. The results are interesting for systems that require a strong LOS component, or systems where a strong LOS component may occur, e.g. for some PTP networks or close to the BS for PMP networks (examples of systems agreeing with this description are given in Section 5).

The analysis is restricted to systems employing ULAs at both sides of the communication link. Some work on this topic has also been done by other authors, e.g. in [52, 93–96]. However, this paper extends this work by i) allowing the arrays to have arbitrary orientation in space[6], ii) characterizing how non-optimal design affects the system parameters, iii) giving analytical expressions for the eigenvalues of the pure LOS channel matrix, and iv) giving an analysis of the approximation introduced to achieve the analytical results.

The MIMO transmission is modeled by the slowly varying frequency-flat fading model given in (1.1). Furthermore, the LOS component is characterized by a simple deterministic RT model, i.e. the LOS model takes into account only the direct components between the Tx and Rx. To evaluate the performance, the LOS channel matrix is employed in the analytical Ricean model from (1.4), where the NLOS part can be correlated and modeled by the Kronecker model. Performance is characterized with respect to the average MI and the MI CDF, both when employing EP allocation and WF power allocation at the Tx.

The optimal design for the pure LOS channel with respect to MI results in orthogonal columns (rows) of $\mathbf{H}_{\mathrm{LOS}}$ when $M > N$ ($M \leq N$). The derivation further reveals that orthogonality is reached when the relation between the antenna separation at Tx and Rx, transmission distance, wavelength, MIMO dimension, and array orientation fulfill a certain criterion. Since the optimal relation is dependent on the Tx–Rx distance and the orientation, designing a system based on this result is best suited for fixed systems. To investigate what happens when the design deviates from this optimal relation, we introduce what we refer to as a deviation factor. We show that even with some deviation from the optimal relation, the potential performance of a pure LOS channel is better than that achieved for an uncorrelated Rayleigh channel (often viewed as a benchmark in MIMO performance analysis). The analytical results derived are

[6] After we first presented our results on ULAs with arbitrary orientation in March 2005 [92], the same results have been reported by another research group in August 2006 [97].

supported by the real world measurements presented in [58, 97–100].

To be able to find the analytical results presented in Paper A, we did perform some approximations. An analysis of these approximations shows that the error introduced increases when; the array size increases, the distance between the Tx and Rx is reduced, and when the array orientation moves away from parallel arrays. A framework on how to analyse this approximation error is presented, and an example system is investigated.

7.2 Paper B

Frode Bøhagen, Pål Orten, and Geir E. Øien, "Optimal Design of Uniform Planar Antenna Arrays for Strong Line-of-Sight MIMO Channels," submitted to *EURASIP Journal on Wireless Communications and Networking, Special Issue on Smart Antennas for Next-Generation Wireless Systems*, November 2006.

In Paper B, which is an extension of the work we presented in [101], we include another dimension in the optimal antenna array design for pure LOS MIMO channels compared to the case described in Paper A, by allowing for planar antenna arrays. More precisely, the antennas are placed in UPAs, with uniform separation in two orthogonal directions forming a lattice of antennas. Some work on the optimal design of UPAs for LOS MIMO channels is presented in [102], however Paper B extends this work by i) allowing for arbitrary orientation of the two arrays (assumed parallel in previous work), ii) including the ULA as a special case, i.e. the optimal design for a communication link with an ULA at one side and an UPA at the other is investigated, and iii) giving analytical expressions for the eigenvalues of the pure LOS channel matrix for some special cases.

As for Paper A, only the direct components between the Tx and the Rx are considered for the LOS channel matrix, and thus for the optimal design. A simple deterministic RT channel model is employed to find these LOS components. The model requires explicit expressions for the path lengths between the Tx and Rx antennas, therefore finding a feasible geometrical characterization of the antenna arrays becomes part of the contribution. The orthogonality requirement, which maximizes the MI for the pure LOS channel, results in a design equation that must be fulfilled to obtain optimal design. This equation is discussed and solved for different scenarios containing both UPAs and ULAs. The LOS channel matrix is employed in the Ricean channel model from (1.4), and performance is evaluated with respect to the MI CDF when EP allocation is employed at the Tx.

7.3 Paper C

Frode Bøhagen, Pål Orten, and Geir E. Øien, "On Spherical vs. Plane Wave Modeling of Line-of-Sight MIMO Channels," conditionally accepted for publication in *IEEE Transactions on Communications* (subject to reviewers' and editor's final approval of revised version currently in preparation).

In this paper we investigate for which transmission scenarios we should employ the true *spherical wave model* (SWM) to model the LOS component in a MIMO channel model, and when the more simple approximate *plane wave model* (PWM) gives sufficient modeling accuracy. The analysis is restricted to systems employing ULAs at both sides of the communication link. To relate these channel models to the characterization introduced in Section 3, both the SWM and the PWM can be viewed as physical deterministic models, where the PWM is an approximation of the exact SWM. Paper C combines our contributions from [103] and [104]; the first investigates the special case of a 2×2 system, while the latter discusses the general case, where the MIMO system can have any dimension. Some numerical investigations have been done on this topic in [58]. However, we present explicit analytical expressions for the transition between these two models.

The analysis is based on the eigenvalues of the pure LOS channel matrix. For the approximate PWM the rank of the LOS channel matrix will always be equal to one, while the rank for the SWM is dependent on the geometrical parameters, e.g. the rank approaches one for large transmission distances. However, for short range communication the rank will often be larger than one, consequently the SWM should be applied to achieve sufficient modeling accuracy. Other parameters that influence the rank are frequency (wavelength), antenna array size, and array orientation.

The level of modeling accuracy required is dependent on the mission and application of the channel model. For example, the channel model can be utilized to evaluate different performance measures as MI, BER, SNR, or outage probability. In Paper C, an explicit expression for the transition between the two models is derived based on the underestimation of MI achieved by employing the PWM instead of the SWM. Results show that for some WLAN scenarios the SWM should be employed to avoid too large underestimation of the MI.

7.4 Paper D

Frode Bøhagen, Pål Orten, Geir E. Øien, and Sébastien de la Kethulle de Ryhove, "Exact Capacity Expressions for Dual-Branch Ricean MIMO Channels," conditionally accepted for publication in *IEEE Transactions on Communications* (subject to reviewers' and editor's final approval of revised version currently in preparation).

In this paper we derive exact analytical expressions for the MI *probability density function* (PDF) and MI CDF for a MIMO Ricean channel as described in (1.4), where no correlation is assumed for the NLOS component, i.e. $\mathbf{R}_{\text{NLOS}} = \mathbf{I}_{MN}$. The investigation is restricted to dual branch MIMO systems, meaning $\min(M, N) = 2$. Expressions for the MI PDF and MI CDF for such a channel were also derived in [72], under the assumption that the rank of $\mathbf{H}_{\text{LOS}}$ is equal to one. However, from Paper A and Paper B we know that the rank of the LOS channel matrix certainly may be larger than one. Consequently, in Paper D this restriction is removed, and the new expressions derived are also valid when the rank of $\mathbf{H}_{\text{LOS}}$ is larger than one.

These expressions are a valuable complement to the MI CDF expressions from [66], since the latter were derived under a Gaussian distributed MI approximation. Although the approximation in [66] in general is quite accurate [105], the error which is committed becomes larger as the dimensions of the MIMO system are reduced.

Paper D is based on the work we presented in [106], where the EP allocation case is treated, and it further extends this work by including an analysis of the MI PDF and MI CDF for the case where WF power allocation is employed (not investigated in [66, 72]). The derivation is based on the distribution of the square of the two singular values of $\mathbf{H}$ (the number of non-zero singular values will always be less or equal to two because of the dual branch assumption). These are employed to (1.8) together with the transmission scheme, EP or WF, to find the MI PDF. The MI CDF is then found by proper integration of the MI PDF. The results show that only one integral must be evaluated numerically to get the MI PDF and MI CDF in the EP allocation case, while two integrals must be evaluated numerically in the WF power allocation case.

The expressions are general and valid for any antenna configuration. To evaluate the results we look at an example where ULAs are employed. The results show that the analytical expressions derived agree perfectly with the simulations performed.

7.5 Paper E

Frode Bøhagen, Pål Orten, and Geir E. Øien, "Modeling and Analysis of a 40 GHz MIMO System for Fixed Wireless Access," *Research Report 353*, Department of informatics, University of Oslo, ISBN 82-7368-310-9, ISSN 0806-3036, March 2007. *Available at: http://www.digbib.uio.no/*

Paper E contains a more comprehensive presentation of the results we first presented in [107]. The objective of the paper is to investigate how the design principle discussed in Paper A works when real world phenomena are taken into account. The main difference between Paper E and [107], is that Paper E includes a more detailed description of some of the models applied. We include the effect of path loss, rain attenuation, antenna array size constraints, and interference from co-channel users in the analysis. We acknowledge that there are other phenomena that could be taken into account as well, especially if we want to make the analysis as realistic as possible, such as e.g. scintillation effects and vegetation effects [108]. However, to make the analysis traceable, we do not want to make the model too complex, and therefore we restrict the analysis to the phenomena mentioned above. Other simulation results based on the design principle from Paper A were recently reported in [109, 110], where a ray-tracing technique is applied to investigate the performance of a 5.2 GHz system in an urban transmission environment.

The transmission technique employed in Paper E is eigenmode transmission, i.e. transmitting along the eigenvectors of the channel matrix. Furthermore, QoS based WF is performed on the resulting spatial channels [111]. This algorithm chooses both the power and transmission mode (coding and modulation) employed on each subchannel, which is based on the available transmission modes and a given QoS requirement. For the system investigated the QoS requirement is BER $\leq 10^{-6}$, and it has eight transmission modes implemented based on *low density parity check* (LDPC) coded QAM/PSK modulation.

By introducing array size restrictions, the performance of the MIMO system will be affected when the transmission distance becomes longer than what allows for optimal design. This is because the length of the optimal ULAs derived in Paper A increases with transmission distance. Consequently, when the distance becomes so large that the maximum array size is reached, the performance will deviate from the optimal design case for longer distances.

The rain has three different effects on the MIMO performance. First of all, when the rain intensity is increased, the attenuation on the desired signal is increased, which has a negative impact on system performance. However, at the same time it reduces the power received from the interfering signals, which is positive. Increased rain intensity also decreases the Ricean K-factor, i.e. the channel becomes more stochastic.

8 Main contributions of the thesis

The main contributions of the thesis can now be summarized to be:

- Finding the optimal design relation with respect to MI for the geometrical parameters for a pure LOS MIMO channel when employing a combination of ULAs and UPAs at the Tx and Rx.
- Characterizing the performance of a LOS MIMO system employing ULAs/UPAs when the design deviates from the optimal. For the ULA case this is done by employing one parameter, i.e. the deviation factor.
- Finding analytical expressions for the eigenvalues for the pure LOS channel matrix when employing various combinations of ULAs and UPAs at the Tx and Rx.
- Describing a technique to decide which LOS channel model that should be employed, i.e. the exact SWM or the approximate PWM, to model the MIMO channel matrix for communication with ULAs. The decision is based on well known performance parameters, such as for example the MI.
- Giving exact analytical expressions for the MI PDF and the MI CDF for a dual MIMO system transmitting over a Ricean channel, both when EP allocation and WF power allocation are employed at the Tx.

- Illustrating how real world effects such as rain and array length constraints will affect the performance of a possible future high frequency fixed wireless access MIMO system employing ULAs.

9 Suggestions for future research

In this section we list some topics that can be interesting to investigate in the future:

- The optimal design for ULAs and UPAs could be employed to a channel model with explicit modeling of multipath (not Rayleigh model). For example, it would be interesting to analyse the performance of such a design when employing a physical deterministic channel model with an environment database. Some work on this is performed in [109, 110], but more scenarios should be investigated.
- The optimal design could be explored for other geometries than ULAs and UPAs. The design could for example be constrained, i.e. the designer has a given area available, and the problem is how should the antennas optimally be placed?
- Based on the results from Paper D, what is the optimal design of the antenna arrays with respect to MI when $K \neq \infty$, i.e. when the channel matrix is not a pure LOS matrix?
- The model decision principle, i.e. SWM versus PWM, for ULAs could be extended to include UPAs as well. This is done by employing the results presented in Paper B.
- Is it possible to exploit knowledge of the deviation factor at the Tx, e.g. investigate different transmission strategies for Ricean channels with deviation factor feedback.

10 Journal and conference contributions during PhD studies

During the PhD studies, the author has contributed to the following journal and conference publications:

- F. Bøhagen, P. Orten, and G. E. Øien, "*Design of Capacity-Optimal High-Rank Line-of-Sight MIMO Channels,*" IEEE Transactions on Wireless Communications, vol. 6, no. 4, April 2007.

- F. Bøhagen, P. Orten, and G. E. Øien, "*On Spherical vs. Plane Wave Modeling of Line-of-Sight MIMO Channels,*" conditionally accepted for publication in IEEE Transactions on Communications.
- F. Bøhagen, P. Orten, G. E. Øien, and S. de la Kethulle de Ryhove, "*Exact Capacity Expressions for Dual-Branch Ricean MIMO Channels,*" conditionally accepted for publication in IEEE Transactions on Communications.
- S. de la Kethulle de Ryhove, G. E. Øien, and F. Bøhagen, "*On the Statistics and Spectral Efficiency of Dual-Branch MIMO Systems with Link Adaptation and Power Control,*" submitted to IEEE Transactions on Vehicular Technology, November 2006.
- F. Bøhagen, P. Orten, and G. E. Øien, "*Optimal Design of Uniform Planar Antenna Arrays for Strong Line-of-Sight MIMO Channels,*" submitted to EURASIP Journal on Wireless Communications and Networking, November 2006.
- F. Bøhagen, P. Orten, and G. E. Øien, "*On Spherical vs. Plane Wave Modeling of Line-of-Sight MIMO Channels,*" in Proc. IEEE Global Communications Conference, San Francisco, USA, November 2006.
- F. Bøhagen, P. Orten, and G. E. Øien, "*Modeling of Line-of-Sight 2x2 MIMO Channels: Spherical versus Plane Waves,*" in Proc. IEEE International Symposium on Personal, Indoor and Mobile Radio Communications, Helsinki, Finland, September 2006.
- F. Bøhagen, P. Orten, and G. E. Øien, "*Optimal Design of Uniform Planar Antenna Arrays for Strong Line-of-Sight MIMO Channels,*" in Proc. IEEE Workshop on Signal Processing Advances in Wireless Communications, Cannes, France, July 2006.
- F. Bøhagen, P. Orten, and G. E. Øien, "*On the Shannon Capacity of Dual MIMO Systems in Ricean Fading,*" in Proc. IEEE Workshop on Signal Processing Advances in Wireless Communications, New York, USA, June 2005.
- F. Bøhagen, P. Orten, and G. E. Øien, "*Modeling and Analysis of a 40 GHz MIMO System for Fixed Wireless Access,*" in Proc. IEEE Vehicular Technology Conference, Stockholm, Sweden, June 2005.
- S. de la Kethulle de Ryhove, Geir E. Øien, and Frode Bøhagen, "*Subchannel SNR Distributions in Dual-Branch MIMO Systems,*" in Proc. IEEE/ITG International Workshop on Smart Antennas, Duisburg, Germany, April 2005.
- F. Bøhagen, P. Orten, and G. E. Øien, "*Construction and Capacity*

Analysis of High-Rank Line-of-Sight MIMO Channels," in Proc. IEEE Wireless Communications and Networking Conference, New Orleans, USA, March 2005.

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